Phase-Only Transmit Beamforming for Spectrum Sharing Microwave Systems

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***Abstract*—This paper deals with the problem of phase-only transmit beamforming in spectrum sharing microwave systems. In contrast to sub-6 GHz schemes, general microwave systems require a large number of antennas due to its huge path loss. As a consequence, digital beamforming needs a large number of computational resources compared to analog beamforming, which only needs a single radio-frequency chain, results the less computational demanding solution. Analog schemes are usually composed by a phase shifter network whose elements transmit at a certain fixed power so that the system designer shall compute the phase values for each element given a set of directions. This approach leads to non-convex quadratic problems where the traditional semidefinite relaxation fails to deliver satisfactory outcomes. In order to solve this, we propose a non-smooth method that behaves well in several scenarios. Numerical evaluations in different spectrum sharing scenarios, which show the performance of our method, are provided.**

I. Introduction

Due to the exponential increase of the traffic demands, not only the cellular wireless access shall be reconsidered but also the backhaul schemes. So far, most of the cellular base stations are connected to the backbone through a digital subscriber line (DSL) connection or, eventually, through an optical fibre link. On the contrary, rural or suburban areas base stations are generally connected via a fixed wireless radio link. These current backhaul approaches suffer from certain disadvantages.

Even though optical fibre links offer an ideally unlimited bandwidth connection, their implantation is costly and its average deployment time is large [1]. Consequently, the resulting capital expenditures (CAPEX) are high. On the other hand, wireless backhaul links cannot offer an unlimited bandwith connection. Nonetheless, they can offer a substantially lower CAPEX and very short deployment time. As a result, wireless backhaul links are of great interest in next generation macro and small cell deployments in both high and low populated areas.

As wireless backhaul links will require a very large bandwidth, both academia and industry are proposing to shift the current microwave radio links to millimeter wave carriers such as the the unlicensed 60 GHz band. Although these frequency bands offer a huge available bandwidth, their path loss and atmospheric degradation effects convert the communication over these carriers a very challenging problem. This is not the case of microwave links in Ka band whose reliability has been tested in the recent years in current deployments.

For both next generation millimetre and microwave backhaul techniques smart antenna techniques are mandatory. Indeed, in contrast to current fixed wireless links, future deployments are expected to be flexible to traffic demands so that the beam pointing reconfigurations are essential. In addition, in case a spectrum sharing scenario is considered (i.e. several communication links share time and frequency resources) interference mitigation techniques are required. This is the case of the deployments in the Ka band where certain satellite receivers could simultaneously operate [2]. In this context, beamforming techniques play a central role.

In contrast to below 3 GHz beamforming techniques, where the spatial processing is generally done in the digital domain, microwave and millimetre wave beamforming techniques require certain processing in the analog domain [3]. This is due to the large number of required radio frequency chains (dozens in microwave and hundreds in millimetre wave), whose all digital processing becomes a cumbersome task. In order to solve this, the system designer could conceive an hybrid design where an analog subsystem transforms the M transmit signals to Q signal to be radiated such as

$$M < Q, \tag{1}$$

so that the digital processing complexity can be drastically reduced. This paper deals with the case where $M = 1$; this is, the spatial processing is all done in the analog domain. Unfortunately, these beamforming techniques show some additional challenges compared to the all digital case.

Analog beamforming relies on a network of phase shifters and power amplifiers whose transmit power is fixed to a certain value leading to the well-known phased array scheme. The optimization of beamforming techniques in phased array structure is an old problem whose convex approximation approach has been investigated in [4]–[6]. As a general statement, obtaining arbitrary complex array beam patterns is a computationally demanding operation which requires genetic algorithms [7] or requires inefficient approximations [6].

Furthermore, as it happens in all digital schemes, transmit beamforming suffers from a large communication overhead due to its required feedback. A limited feedback analog beamforming scheme can be found in [8] where codebook and access schemes are presented. Indeed, this joint beamforming and access approach is a key challenge also in microwave

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systems, where the acquisition time requires that each beam (transmitter and receiver) scans in all the angle coverage. Another example of this can be found in [9]. Yet another approach is to consider during the scanning period a channel estimation (direction of arrival) based on compress sensing techniques as in [10]. In light of the aforementioned papers, in this preliminary work we consider that the transmitter has access to the interfering direction leading to the so-called spatial reference beamforming.

In contrast to previously exposed works, this paper presents a general optimization framework for phase-only transmit beamforming designs in spectrum sharing systems. Considering that the desired and interfering directions are known by the transmitter, we propose an efficient optimization method.

The design of the optimal beamforming under the aforementioned conditions leads to a non-convex quadratically constrained quadratic program. The typical approach for solving such problems is to use a semidefinite relaxation (SDR) followed by a Gaussian randomization technique [11]. Nonetheless, under the presence of individual power constraints, which are needed for the design of a phase-only beamforming, such approach fails to deliver a satisfactory outcome. In this paper, we trackle the aforementioned optimal beamforming design as a semidefinite programming (SDP) with an additional reverse (but continuous) constraint. Non- smooth optimization algorithms are then proposed to locate the optimal solutions of such design problem. Similar approaches have been considered before in other scenarios [12], [13].

The proposed scheme is evaluated considering a spectrum sharing scenario where both a wireless backhaul link takes place in presence of non-intended receivers. Remarkably, the proposed scheme is able to control the interfere power levels in different directions which is adequate for next generation spectrum sharing systems in contrast to other techniques such as [14].

The paper is organized as follows. Section II introduces the system. Section III formulates the optimization problem and the relaxation one with ideal phase quantization and one bit quantization. Section IV proposes an optimization technique for improving the semidefinite relaxation method. Section VI illustrates the performance of the proposed techniques and Section VII concludes.

Notation: We adopt the notation of using lower case boldface for vectors, $\mathbf{v}$, and upper case boldface for matrices, $\mathbf{A}$. The transpose operator and the conjugate transpose operator are denoted by the symbols $(\cdot)^T$, $(\cdot)^H$ respectively. $\mathbf{I}$ denotes the identity matrix. $\mathbb{C}$ denotes the complex numbers. $\|\cdot\|$ denotes the Euclidean norm. $|\cdot|$ denotes the absolute value. $\circ$ denotes the Hadamard product.

II. System Model

Let us consider a base station equipped with N antennas transmitting a unit energy symbol s to a certain receiver. The received signal can be modelled as

$$y = \sqrt{PG}\mathbf{a}(\theta)^H \mathbf{w}_{\text{ULA}} s + n, \tag{2}$$

where $\mathbf{a}(\theta) \in \mathbb{C}^{Q\times 1}$ is the array antenna response that depends on the angle of departure (AoD) between the transmitter and the receiver ($\theta \in [-\pi, \pi]$) and the array element structure. For the sake of simplicity, we will consider an uniform linear array (ULA) whose array antenna response is

$$\mathbf{a}_{\text{ULA}}(\theta) = \frac{1}{\sqrt{Q}}\left(1, e^{j\frac{2\pi}{\lambda}d\sin(\theta)}, \dots, e^{j\frac{2\pi}{\lambda}(Q-1)d\sin(\theta)}\right)^T, \tag{3}$$

where d is the antenna spacing and λ the transmission wavelength. The path-loss is modelled by G and P denotes the transmit power. Vector $\mathbf{w}_{\text{ULA}} \in \mathbb{C}^{Q\times 1}$ denotes the beamforming vector to be designed and n the Gaussian distributed zero mean unit variance noise term.

Microwave backhaul links generally operate with planar arrays due to its high directivity. The simplest planar array representation is a uniform rectangular array (URA). These arrays are usually represented in matrix form but, it is more convenient we consider its vector formulation as follows

$$\mathbf{a}_{\text{URA}}(\theta,\phi) = \text{vec}\left(\mathbf{u}(\theta,\phi)\,\mathbf{v}(\theta,\phi)^T\right), \tag{4}$$

where $\mathbf{u}(\theta,\phi)$ and $\mathbf{v}(\theta,\phi)$ are described in (5) and (6).

Parameters d_x and d_y are the antenna distances in the x and y axis and N_x and N_y are the number of elements in the x and y axis respectively. The azimuth angle is represented by $\phi \in [0, \pi]$. Under this context, the system designer shall optimize the $N_xN_y = Q$ complex vector $\mathbf{w}_{\text{URA}} \in \mathbb{C}^{Q\times 1}$ in order to obtain a reliable communication.

In addition, there might be case where an arbitrary antenna array is used. In that case, the antenna array response can be represented by

$$\mathbf{a}_{\text{ARB}} = \mathbf{a}_{\text{URA}}(\theta,\phi) \circ \mathbf{e}, \tag{7}$$

where $\mathbf{e}$ is $Q \times 1$ vector whose elements are 1 and 0 whenever the antenna element is present or not respectively.

Apart from the intended user, this paper considers that the transmission takes place in presence of K non-intended users sharing time and spectral resources. For instance, these can be satellite receivers aiming to detect the information shared by the satellite transmission. In this scenario, terrestrial backhaul links operating at the same frequency can eventually create an interference signal so that the satellite receivers are unable to establish a reliable communication link with the satellite.

These interfered users are modelled by a set of AoDs $\left(\{(\theta_k,\phi_k)\}_{k=1}^{K-1}\right)$, leading to the following array antenna responses

$$\{\mathbf{a}_i^k(\theta_k,\phi_k)\}_{k=1}^{K}. \tag{8}$$

The corresponding received signals can be modelled as

$$y_k = \sqrt{PG_k}\mathbf{a}_k(\theta_k,\phi_k)^H \mathbf{w} s + n_k \quad k = 1,\dots,K-1, \tag{9}$$

where G_k is the path loss for the k non-intended receiver and n_k the Gaussian distributed zero mean unit variance noise term.

This paper focuses on the design of $\mathbf{w}$[1] so that it maximizes the communication rate with the intended receiver while keeps the interference power levels to the interfered users under certain threshold.

[1]We now skip the subindex URA, ULA and ARB and we consider an arbitrary array.

$$\mathbf{u}\left(\theta,\phi\right)=\frac{1}{\sqrt{N_x}}\left(1,e^{j\frac{2\pi}{\lambda}d_x\sin(\theta)\cos(\phi)},\ldots,e^{j\frac{2\pi}{\lambda}(N_x-1)d_x\sin(\theta)\cos(\phi)}\right)^T \tag{5}$$

$$\mathbf{v}\left(\theta,\phi\right)=\frac{1}{\sqrt{N_y}}\left(1,e^{j\frac{2\pi}{\lambda}d_y\sin(\theta)\sin(\phi)},\ldots,e^{j\frac{2\pi}{\lambda}(N_y-1)d_y\sin(\theta)\sin(\phi)}\right)^T \tag{6}$$

Additionally, in contrast to previous works, it is considered that the beamforming is performed in the analog domain by means of a set of Q phase shifters. Under this context, the beamforming vector shall be constrained so that

$$|[\mathbf{w}]_i|^2=\frac{1}{Q}\quad i=1,\ldots,Q. \tag{10}$$

The following sections describe different optimizations of the beamforming vector considering different spectrum sharing scenarios.

III. Phase-Only Beamforming Optimization

A. Unicast Transmission

Let us consider a transmit beamforming optimization in presence of K non-intended receivers. Whenever the path losses and the AoDs are available, the system designer shall optimize the following problem

$$\begin{aligned}&\underset{\mathbf{w}}{\text{maximize}}\quad |\mathbf{a}_d^H\mathbf{w}|^2\\&\text{subject to}\\&|\mathbf{a}_i^{k,H}\mathbf{w}|^2\leq\epsilon_k\quad k=1,\ldots,K-1,\\&|[\mathbf{w}]_i|^2=1/Q\quad i=1,\ldots,Q,\end{aligned} \tag{11}$$

where ϵ_k for $k=1,\ldots,K$ denote the maximum array gain to the non-intended receivers. Moreover, $\mathbf{a}_d$ denotes the AoD array antenna response to the intended user.

The optimization problem in (11) is non-convex quadratically constraint quadratic program due to the equality constraint.

B. Multicast Transmission

Whenever a group of users want to receive the same content, multicast transmissions can substantially increase the spectral efficiency. With this, the transmitter must ensure that a certain symbol is decoded by all users. Under this context, the achievable rate is dictated by the user with lowest SNR, leading to the following optimization problem

$$\begin{aligned}&\underset{\mathbf{w}}{\text{maximize}}\quad \underset{l=1,\ldots,D}{\text{minimum}}\quad |\mathbf{a}_d^{(l,H)}\mathbf{w}|^2\\&\text{subject to}\\&|\mathbf{a}_i^{H,k}\mathbf{w}|^2\leq\epsilon_k\quad k=1,\ldots,K-1,\\&|[\mathbf{w}]_i|^2=1/Q\quad i=1,\ldots,Q,\end{aligned} \tag{12}$$

where we consider a set of D intended users whose array antenna responses $\{\mathbf{a}_d^l\}_{l=1}^L$. As for the unicast optimization (12) is a non-convex QCQP. In any case, it is more convenient to rewrite it as

$$\begin{aligned}&\underset{\mathbf{w}}{\text{maximize}}\quad t\\&\text{subject to}\\&|\mathbf{a}_i^{k,H}\mathbf{w}|^2\leq\epsilon_k\quad k=1,\ldots,K,\\&|\mathbf{a}_d^{l,H}\mathbf{w}|^2\geq t\quad l=1\ldots,L,\\&|[\mathbf{w}]_i|^2=1/Q\quad i=1,\ldots,Q,\end{aligned} \tag{13}$$

Under this context, we can observe that both (13) and (11) are equivalent. Indeed, the unicast transmission sets $L=1$. As a result, we will consider this later case as it collapses also the unicast optimization problem.

IV. Non-convex QCQP Optimization

In order to obtain an efficient solution of (13) optimization problem, we propose two different approaches; namely, the semidefinite relaxation and the non-smooth optimization. As it is described, even though the semidefinite relaxation technique is the most popular approximation for solving QCQP optimization problems, it fails for this optimization problem. As a result, the non-smooth optimization problem becomes the mandatory solution.

A. Semidefinite Relaxation

The semidefinite relaxation technique relays on dropping the rank-one restriction. With this, the optimization problem (13) can be re-written as

$$\begin{aligned}&\underset{\mathbf{w}}{\text{maximize}}\quad t\\&\text{subject to}\\&\text{Tr}\left(\mathbf{A}_i^{H,k}\mathbf{W}\right)\leq\epsilon_k\quad k=1,\ldots,K,\\&\text{Tr}\left(\mathbf{A}_d^{H,l}\mathbf{W}\right)\geq t\quad d=1,\ldots,D,\\&\text{diag}\left(\mathbf{W}\right)=\frac{1}{Q}\mathbf{1},\end{aligned} \tag{14}$$

where

$$\mathbf{A}_i=\mathbf{a}_i^k\mathbf{a}_i^{k,H}\quad k=1,\ldots,K, \tag{15}$$

$$\mathbf{A}_d=\mathbf{a}_d^k\mathbf{a}_d^{l,H}\quad l=1,\ldots,L. \tag{16}$$

Under this context, whenever the optimal solution of (14), $\mathbf{W}^*$ is rank one, the original optimization problem (13) has an optimization solution $\mathbf{w}^*$ so that

$$\mathbf{W}^*=\mathbf{w}^*\mathbf{w}^{H,*}. \tag{17}$$

Unfortunately, this tight approximation does not generally occurs and $\mathbf{W}^*$ generally yields to a high rank solution.

In order to solve this problem, Gaussian randomization technique is used. This technique relies on the computation of vector Gaussian realizations so that

$$\mathbf{w}_g\sim\mathcal{N}\left(\mathbf{0},\mathbf{W}^*\right). \tag{18}$$

After that, the Gaussian randomization, $\mathbf{w}_g$ needs to be adapted in order to attain the constraints. This adaptation is problem-dependent and for our case it becomes

$$\mathbf{w}_g \leftarrow \mathbf{G} \circ \mathbf{w}_g, \tag{19}$$

where

$$\mathbf{G} = \operatorname{diag}\left(\frac{1}{|[\mathbf{w}_g]_1|}, \dots, \frac{1}{|[\mathbf{w}_g]_Q|}\right). \tag{20}$$

Note that with the operation (19), the randomization, $\mathbf{w}_g$, becomes a phase-only scheme. Once this operation is done, the interference restrictions constraints are evaluated

$$|\mathbf{a}_i^{k,H}\mathbf{w}_g|^2 \leq \epsilon_k \quad k = 1, \dots, K, \tag{21}$$

and, in case they are fulfilled, the randomization $\mathbf{w}_g$ is considered as a valid solution.

The process is done several times $(10^3, 10^4)$ and the solution that yields the maximum objective function is chosen. In the following we summarize the algorithm.

Data: $\mathbf{W}^*$
Result: $\mathbf{w}_g^*$
initialization;
for $n = 1$ *to* $n = N_{\textit{randomizations}}$ **do**
  Generate $\mathbf{w}_g \sim \mathcal{N}(\mathbf{0}, \mathbf{W}^*)$;
  $\mathbf{w}_g \leftarrow \mathbf{G} \circ \mathbf{w}_g$;
  if $\mathbf{w}_g$ *fulfils* (21) **then**
    save $\underset{l=1,\dots,L}{\text{minimum}} \quad |\mathbf{a}_d^{(l,H)}\mathbf{w}|^2$;
  else
    go back and generate another randomization;
  end
end
Elect the randomization $\mathbf{w}_g$ which has delivered the largest $\underset{l=1,\dots,L}{\text{minimum}} \quad |\mathbf{a}_d^{(l,H)}\mathbf{w}|^2$ value ;

Algorithm 1: Gaussian randomization techniques for phase-only spectrum sharing transmit beamforming techniques

As it is shown in the numerical simulations section, it occurs that even through an extremely high value of $N_{\text{randomizations}}$ is chosen, the aforementioned algorithm does not lead to any valid solution. This is due to adaptation technique (19) that it does not take into account the interference restrictions values.

B. Nonsmooth Optimization

Since the semidefinite relaxation technique does not offer an efficient solution we propose in the following an alternative optimization technique.

As discussed in [12], [13], the inefficient semidefinite relaxation technique generally fails due to the discontinuous nature of the rank one constraint. A more natural way of impossing the rank-one restrictions in semidefinite positive matrices is to write it as

$$\operatorname{Tr}(\mathbf{W}) - \lambda_{\text{max}}(\mathbf{W}) \leq 0, \tag{22}$$

where $\lambda_{\text{max}}(\mathbf{W})$ refers to the maximum eigenvalue of matrix $\mathbf{W}$. It can be observed that indeed, (22) is equivalent to

$$\operatorname{rank}(\mathbf{W}) = 1, \tag{23}$$

since $\operatorname{Tr}(\mathbf{W}) = \lambda_{\text{max}}(\mathbf{W})$ whenever there is only one non-zero eigenvalue.

Therefore, the optimization problem (14) can incorporate the rank-one constraint as follows

$$\begin{aligned}
&\underset{\mathbf{W}}{\text{maximize}} \quad t\\
&\text{subject to}\\
&\operatorname{Tr}\left(\mathbf{A}_i^{H,k}\mathbf{W}\right) \leq \epsilon_k \quad k = 1, \dots, K,\\
&\operatorname{Tr}\left(\mathbf{A}_d^{H,l}\mathbf{W}\right) \geq t \quad d = 1, \dots, L,\\
&\operatorname{diag}(\mathbf{W}) = \frac{1}{Q}\mathbf{1},\\
&\operatorname{Tr}(\mathbf{W}) - \lambda_{\text{max}}(\mathbf{W}) \leq 0,
\end{aligned} \tag{24}$$

We take the approach of the penalty function so that we incorporate the new constraint into the objective function

$$\begin{aligned}
&\underset{\mathbf{W}}{\text{maximize}} \quad t - \mu\left(\operatorname{Tr}(\mathbf{W}) - \lambda_{\text{max}}(\mathbf{W})\right)\\
&\text{subject to}\\
&\operatorname{Tr}\left(\mathbf{A}_i^{H,k}\mathbf{W}\right) \leq \epsilon_k \quad k = 1, \dots, K,\\
&\operatorname{Tr}\left(\mathbf{A}_d^{H,l}\mathbf{W}\right) \geq t \quad d = 1, \dots, L,\\
&\operatorname{diag}(\mathbf{W}) = \frac{1}{Q}\mathbf{1}.
\end{aligned} \tag{25}$$

Even though problem (25) is convex, the function $\lambda_{\text{max}}(\mathbf{W})$ is not differentiable. In order to solve this problem, a sub-differential of the largest eigenvalue function is used instead. This sub-differential version of the function is

$$\partial\lambda_{\text{max}}(\mathbf{W}) = \mathbf{w}_{\text{max}}\mathbf{w}_{\text{max}}^H, \tag{26}$$

where $\mathbf{w}_{\text{max}}$ is the eigenvector associated to the largest eigenvalue. Bearing in mind that

$$\lambda_{\text{max}}(\mathbf{Y}) - \lambda_{\text{max}}(\mathbf{W}) \geq \operatorname{Tr}\left(\mathbf{w}_{\text{max}}\mathbf{w}_{\text{max}}^H(\mathbf{Y} - \mathbf{X})\right), \tag{27}$$

for any semidefinite positive matrix $\mathbf{Y}$, we can construct an iterative algorithm for obtaining an efficient solution to (25).

Indeed, given an optimal solution of problem (14) $\mathbf{W}^*$ with its corresponding maximum eigenvalue $\lambda_{\text{max}}(\mathbf{W}^*)$ optimization problem (28) provides a better solution.

Consequently, we can iteratively compute the optimal solution of (28) based on a previous optimal solution. Note that it is essential that a feasible initial solution $\mathbf{W}^{(0)}$ is found so as the value μ is properly elected. While the initial feasible solution can be obtained from (14), the election μ depends on the problem and it shall be updated in case the algorithm remains in a high rank solution. In the following we summarize this methods.

V. Numerical Results

In this section we evaluate the performance of the proposed technique in a close-to-real simulation. We will consider that a base station aims at transmitting its aggregated traffic towards one (unicast) or more (multicast) base stations. The 18 GHz band is employed in presence of other non-intended user which can be either satellite receivers or other wireless backhaul links.

$$\begin{aligned}
&\underset{\mathbf{W}}{\text{maximize}} \quad t-\mu\left(\text{Tr}\left(\mathbf{W}\right)-\lambda_{\text{max}}\left(\mathbf{W}^{*}\right)-\text{Tr}\left(\mathbf{w}_{\text{max}}\mathbf{w}_{\text{max}}^{H}\left(\mathbf{W}-\mathbf{W}^{*}\right)\right)\right)\\
&\text{subject to}\\
&\text{Tr}\left(\mathbf{A}_{i}^{H,k}\mathbf{W}\right)\leq\epsilon_{k}\quad k=1,\ldots,K,\\
&\text{Tr}\left(\mathbf{A}_{d}^{H,l}\mathbf{W}\right)\geq t\quad d=1,\ldots,L,\\
&\text{diag}\left(\mathbf{W}\right)=\frac{1}{Q}\mathbf{1}.
\end{aligned} \tag{28}$$

Data: $\mathbf{W}^{(0)}$ from (14) and μ
Result: $\mathbf{w}^{*}$
initialization ;
while $\mathbf{W}^{(n)}$ *is not rank one* **do**
  Compute $\mathbf{W}^{(n+1)}$ according to (28).;
  if $\mathbf{W}^{(n+1)}=\mathbf{W}^{(n)}$ **then**
    Update μ so that $\mu\leftarrow 2\mu$;
  else
    $n\leftarrow n+1$;
    $\mathbf{W}^{(n+1)}\leftarrow\mathbf{W}^{(n)}$;
  end
end
Output the final solution;

Algorithm 2: Non-smooth optimization for phase-only spectrum sharing transmit beamforming techniques

For an easy comparison, we will consider an operative point fixed 19 dBm transmit power so that the desired receivers are situated at 1 Km distance. In addition, clear sky conditions are assumed so that only the path loss effect is taken into account.

It is considered that the interfered users uniformly located over the AoDs and the transmitter shall reduce the array gain in those directions a uniform random value from -60 and -40 dBs. In all cases, we consider a total number of 1000 realizations. The next table summarizes the simulation parameters where the ETHERNET ©radio transceiver has been considered [2].

TABLE I. USER LINK PARAMETERS

Parameter	Value
Distance	1 km
User location distribution	Uniformly distributed
Carrier frequency	18 GHz (Ka band)
Atmospheric fading	Clear Sky
Number of antenna elements	25
Antenna element gain	10 dB
Bandwidth	56 MHz
Modulation	64 QAM
Sensitivity for 10^{-6} BER	- 74 dBm
Transmit power	19 dBm

We compare our phase-only technique with the all-digital technique. Mathematically, we compare our method with the optimal solution of

$$\begin{aligned}
&\underset{\mathbf{w}}{\text{maximize}} \quad t\\
&\text{subject to}\\
&\text{Tr}\left(\mathbf{A}_{i}^{H,k}\mathbf{W}\right)\leq\epsilon_{k}\quad k=1,\ldots,K,\\
&\text{Tr}\left(\mathbf{A}_{d}^{H,l}\mathbf{W}\right)\geq t\quad d=1,\ldots,L,\\
&\text{Tr}\left(\mathbf{W}\right)\leq 1.
\end{aligned} \tag{29}$$

Note that for both cases the same transmit available power is considered. Furthermore, the cumulative distributed function (CDF) of the data rate is computed in order to obtain a more complete comparison of all methods. The rate for the unicast case is defined as

$$R_{\text{unicast}}=B\log_{2}\left(1+\frac{PG|\mathbf{a}_{d}^{H}|^{2}}{\sigma^{2}}\right), \tag{30}$$

where B is the user bandwidth. For the multicast case, the rate is obtained

$$R_{\text{multicast}}=\underset{l=1,\ldots,D}{\text{minimum}}\quad B\log_{2}\left(1+\frac{PG|\mathbf{a}_{d}^{l,H}|^{2}}{\sigma^{2}}\right). \tag{31}$$

In Figure 1 a URA is considered with 25 elements. Curiously, for this case the semidefinite relaxation always yields to rank-one solution. Therefore, our non-smooth method is not required for this case. The authors think that this is related with the fact that since an URA is represented by a Vandermonde matrix an as in [15], this makes the semidefinite relaxation tight. However, this is not shown in this paper and it is left for further studies.

It can be observed in Figure 1 that even though there exist a degradation between the phase-only scheme with respect to the all-digital, this is minimal.

In order to observe the need of our non-smooth approach, we simulate an arbitrary antenna array randomly deploying antenna elements overall an URA. For this case, the semidefinite relaxation did not deliver any valid solution over the 1000 realizations and considering $N_{\text{randomization}}=10^{6}$. On the contrary, our non-smooth approach always yields a solution and its rate performance is depicted in Figure 2. It is shown that the degradation with respect to the URA case is larger.

For the multicast case we consider a concrete scenario where the transmit beamforming scheme requires to reject the interference of two AoD located at $\theta_{i}^{1}=30,\phi_{i}^{1}=0$ and $\theta_{i}^{2}=-30,\phi_{2}^{1}=60$. The interfering power signals shall be reduced $\epsilon=40$ dB. Additionally, the desired AoDs are located at $\theta_{d}^{1}=45,\phi_{d}^{1}=45$ and $\theta_{d}^{2}=-45,\phi_{d}^{2}=-45$.

[2]https://www.winncom.com/pdf/BridgeWave_EtherFlex/BridgeWave_EtherFlex_DS.pdf

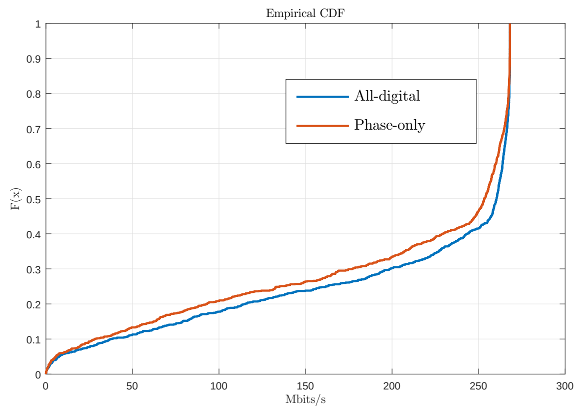

Fig. 1. Rate performance of a transmit beamforming scheme in a URA with 1 interfering user.

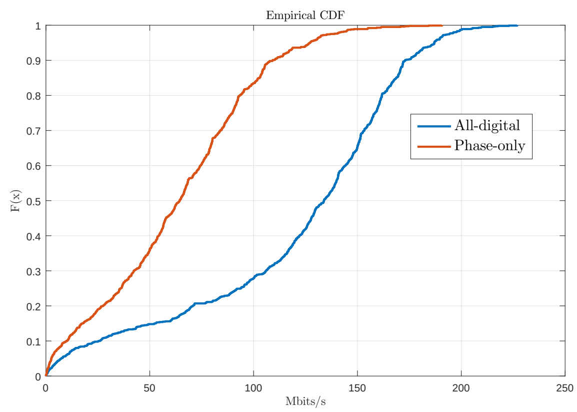

Fig. 2. Rate performance of a transmit beamforming scheme in an arbitrary rectangular array with 1 interfering user.

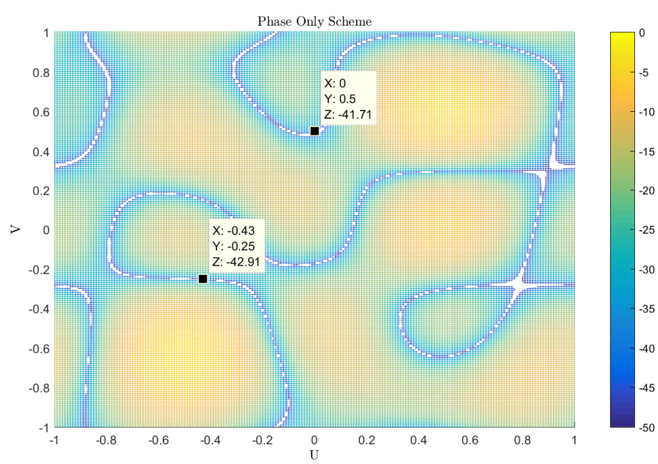

Fig. 3. Phase-only multicast array factor optimized via the non-smooth scheme. The desired AoD are $\theta_d^1 = 45, \phi_d^1 = 45$ and $\theta_d^2 = -45, \phi_d^2 = -45$ whereas the interference AoD are $\theta_i^1 = 30, \phi_i^1 = 0$ and $\theta_i^2 = -30, \phi_i^2 = 60$. Moreover, ϵ is set to -40 dB.

For this case, the semidefinite method yields to a high rank solution so that the non-smooth approach becomes mandatory again even though an URA is considered. It is depicted in figures 3 and 4 that in both cases the optimization method is able to mitigate the interference at the indicated directions at least -40 dBs. The main different between both schemes is the array gain the desired directions. For this example, the all-digital scheme is able to provide an array gain of -3.4 dB in both directions whereas the phase-only scheme can only provide -5.7 dB. Note that the array gain factors are represented in dBs in the U-V plane where $U = \sin(\theta)\cos(\phi)$ and $V = \sin(\theta)\sin(\phi)$.

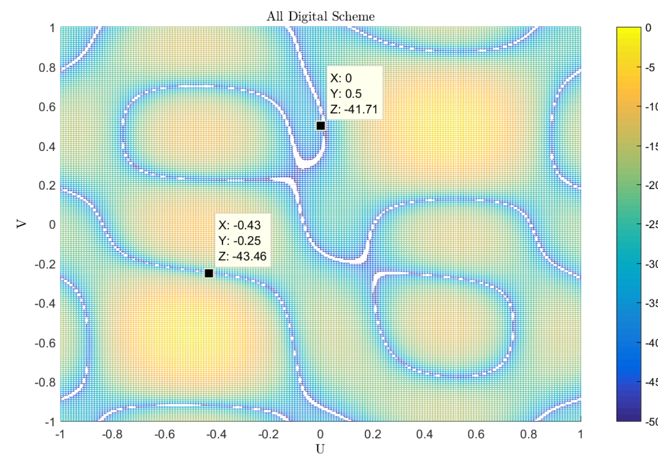

Fig. 4. All-digital multicast array factor optimized via the non-smooth scheme. The desired AoD are $\theta_d^1 = 45, \phi_d^1 = 45$ and $\theta_d^2 = -45, \phi_d^2 = -45$ whereas the interference AoD are $\theta_i^1 = 30, \phi_i^1 = 0$ and $\theta_i^2 = -30, \phi_i^2 = 60$. Moreover, ϵ is set to -40 dB.

Finally we evaluate our approach considering an arbitrary rectangular array and a multicast communication with two desired directions and two interfering ones. In Figure 5 the rates are depicted and it can be observed that the achievable rates are lower than the unicast case. Moreover, the difference between the phase-only scheme and the all-digital one is high as for the unicast case presented in figure 2.

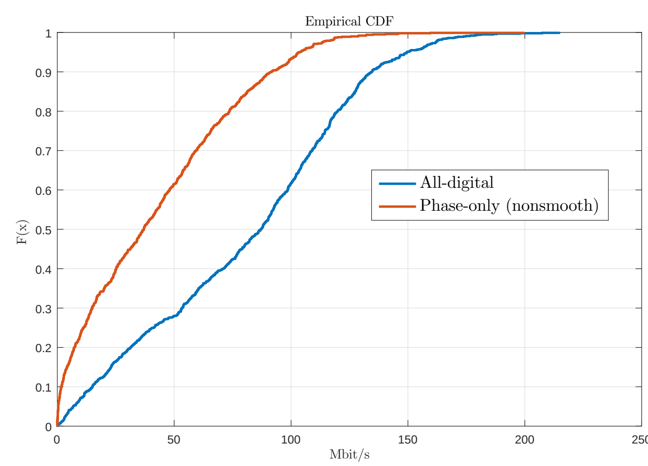

Fig. 5. Rate performance of a transmit beamforming scheme in an arbitrary rectangular array with 2 interfering user in a multicast communication with 2 desired users.

VI. Conclusions

This paper proposes a non-smooth optimization which is adequate for next generation wireless backhaul beamforming. In contrast to all-digital designs, where a baseband receiver requires large computational resources we propose a phase-only analog scheme, which only requires a single radiofrequency chain and a set of digitally controlled phase shifters. Our proposed scheme is the only available tool for efficiently obtaining the beamforming vectors for phase-only spectrum sharing cases as the semidefinite relaxation is not able to deliver a solution even though a high number of randomizations are considered. Despite certain degradation is observed compared to the all-digital scheme, the obtained rates are high, leading to a low complex solution for backhaul links in spectrum sharing environments.

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